Remarks/Arguments

In view of the comments below Applicant respectively requests that the Examiner reconsider the present application, including claims 1-38 and claims 77-83, and withdraw the rejections of claims 1-4, 6-9, 11, 16, 17 21-25, 28, 33-37, 77-79, 82 and 83 and the objections to claims 5, 10, 12-15, 18-20, 26, 27, 29-32, 38, 80 and 81.

a) Claims 1-4, 6-9, 11, 21, 25, 28, 33-37, 77-79, 82 and 83 stand rejected under 35 U.S.C. 102(b) as being anticipated by Tripathi, et al. (U.S. Patent No. 5,974,089), (Fig. 2).

As an overview of the discussion below, Applicant respectfully submits that Tripathi, et al. '089 does not apply to radio frequency amplifiers and further does not show or suggest the claimed switching mode power amplifier … to provide a radio frequency signal or a linearizer or a linearizer intercoupled as claimed or a linearizer using an adaptive process as claimed.

Tripathi et al ('089) concerns an approach for limiting the minimum time between transitions of pulses that are applied to an audio power amplifier 312 (see for example, abstract). More specifically, Fig. 2 shows a standard oversampled low pass second order (integrators 202, 206) sigma delta modulator with an output comparator stage 210 driving a power switch stage 212. Fig. 3 of Tripathi et al '089 shows the diagram of Fig. 2 with an additional pulse qualification logic 318 and the corresponding description of 318 indicates as noted above that the function of logic 318 is to ensure a minimum time period between transitions (col.6, lines 12-14).

Tripathi et al '089 is an arrangement that is applicable to low frequency applications where narrow pulses are unacceptable, such as audio amplifiers ... (col.6, lines 50-54). In short Tripathi et al '089 is not applicable to radio frequency power amplifiers and in fact what is taught by Tripathi et al '089 would render a radio frequency amplifier impossible to construct, since as is known a radio frequency amplifier will necessarily have narrow pulses.

Regarding the Examiner's comments and referring to claims 1-3, 6-9, 11, 21, 28, 33-37, 77-79, 82 and 83, the Examiner maintains that "Tripathi et al (Fig. 2) discloses an amplifier circuit comprising a delta sigma modulator (206, 210) which is connected to receive an input signal (input) and produces a bi-level modulation signal (output of 210), a switching mode power amplifier (212) which is driven by the bi-level modulation signal (output of 210) and having an output (output), and a linearizer (204,202,208,214,216) which is coupled to the input signal (input) and the RF signal (output) to supply a corrective signal (output of 208) at a location prior to the switching mode power amplifier (212) wherein the linearizer (204,202,208,214,216) using an adaptive process."

Applicant respectfully disagrees with the Examiner's construction noting, for example, that the delta sigma modulator 206, 210 (using the Examiner's construction) is not connected to the input signal; rather 206, 210 are connected between the output of a summer 208 and input of a power switch 212 and none of the terminals of 206 or 210 are connected to the input signal. At best it might be argued that 206 is coupled to something somehow corresponding to the input signal, but certainly not the input signal. Furthermore 204, 202, 208, 214, 216 as depicted by Fig. 2 is not a linearizer as in so far as Applicant can determine these blocks do nothing to

linearize a system (e.g. modify operation of a system, such as modify a signal or operational characteristics of a circuit to reduce overall distortion in the composite system). For example, these blocks do not compensate for distortion caused by switching stage 212 or any other stage by modifying an operating point for this stage or canceling or otherwise compensating for distortion or the like. These functional blocks are simply part of the delta sigma modulator of Tripathi et al '089.

Furthermore, this collection of functions is not adaptive in any sense of the word, as all of these stages have fixed functionality. The Examiner seems to be of the view that any collection of symbols in a figure can be appropriately referred to as a linearizer. Applicant strongly disagrees with this casual construction of Tripathi et al '089. For the Examiner's information and convenience, Applicant has enclosed a paper by Grant et al entitled "A DSP Controlled Adaptive Feedforward Amplifier Linearizer" that was published in Vehicular Technology, Vol. 44 No. 1, pgs 31-40 February 1995 and discussed an Adaptive Amplifier linearizer. The second enclosure is selected pages of a 1996 Thesis by Grant entitled "A DSP Controlled Adaptive Feedforward Amplifier Linearizer". The Thesis discusses some linearizer background beginning at page 4 showing that linearizers are known and were known at the time of Applicant's invention. Each of these papers includes a listing of additional references. However it was not known to combine such linearizers with the other structures of the claimed invention. Furthermore, the switching mode power amplifier (power switch 212) of Tripathi et al '089 does not show or suggest the claimed switching mode power amplifier ... operable to provide a radio frequency signal.

Applicant's claim 1 recites:

"A radio frequency amplifier system comprising:

a delta sigma modulator connected to receive an input signal and produce a bi-level modulation signal;

a switching mode power amplifier driven by the bi-level modulation signal and operable to provide a radio frequency signal at an output; and

a linearizer, coupled to the input signal and the radio frequency signal, operable to supply a corrective signal at a location prior to the switching mode power amplifier, the linearizer using an adaptive process.

Tripathi et al 089 does not show or suggest and in fact teaches away from a radio frequency amplifier. Applicant concedes that the reference shows or suggests a delta sigma modulator more or less as claimed. However this reference does not show or suggest the claimed switching mode power amplifier providing a radio frequency signal and the reference does not show or suggest a linearizer or linearizer intercoupled as claimed or a linearizer that is adaptive. Therefore Applicant respectfully submits that Tripathi et al 089 does not support a rejection of claim 1 under §102(b). Claims 2-38 are dependent on claim 1 and at least by virtue of this dependency on a claim that is believed to be allowable should likewise be deemed allowable.

Claim 77 is another independent claim that defines:

"A radio frequency amplifier system comprising:

a bandpass delta sigma modulator connected to receive an input signal and produce a bi-level modulation signal;

a switching mode power amplifier driven by the bi-level modulation signal and having an output; and

a linearizer, coupled to the input signal and the output of the switching mode power amplifier, operable to supply a corrective signal at a location prior to the switching mode power amplifier, the linearizer using an adaptive linearization process."

Tripathi et al '089, for the reasons noted above, does not show or suggest a radio frequency amplifier. Furthermore, Tripathi et al '089 does not show or suggest a bandpass delta sigma modulator as claimed, although Applicant submits that other references included with the IDS may show such a structure. Additionally Tripathi et al '089 as noted above does not show or suggest a linearizer or linearizer intercoupled as claimed or linearizer using adaptive process as claimed. Thus Applicant respectfully submits that Tripathi et al '089 does not support a rejection of claim 77 under §102(b). Claims 77-83 are dependent on claim 77 and at least by virtue of this dependency on a claim that is believed to be allowable should likewise be deemed allowable.

Therefore and at least for these reasons, Applicant respectfully request the Examiner to reconsider and withdraw this rejection of claims 1-4, 6-9, 11, 21, 25, 28, 33-37, 77-79, 82 and 83 under 35 U.S.C. 102(b) based on Tripathi, et al. (U.S. Patent No. 5,974,089), (Fig. 2).

Additional reasons for allowing the dependent claims will now be discussed.

Other limitations of various of the dependent claims are not shown or suggested by Tripathi et al '089 taken alone or together with any other reference of record and thus various of the dependent claims are allowable for additional reasons that will be briefly and at least partially summarized below for the Examiner's convenience.

Regarding claim 3, Tripathi et al does not show or suggest a multi-band bandpass delta sigma modulator. Claim 78 is similar to claim 3 but dependent on claim 77. Thus and for this additional reason, Applicant respectfully submits that Tripathi et al '089 does not show or

suggest the claimed multi-band bandpass delta sigma modulator of claim 3 or 78 and does not support a rejection of claim 3 or 78 under §102(b).

Regarding claim 4, the Examiner maintains that "Tripathi et al. (Fig. 2) inherently includes a tunable output filter since it would not work without the filter." While it may be reasonable to argue that Tripathi needs an output filter, it is clearly erroneous to suggest that a tunable filter is needed. Applicant respectfully notes that Tripathi et al '089 addresses audio frequency applications or like low frequency applications and thus in Applicant's view almost certainly would not use a tunable filter. Applicant also notes that claim 4 recites a tunable output filter tunable to a plurality of frequency bands which clearly is not shown or suggested by this or any other reference of record. Thus and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed tunable output filter of claim 4 and does not support a rejection of claim 4 under §102(b).

Claim 8-9, 11, 16-17, and 35 recite various limitations corresponding to an extended interface between the delta sigma modulator and the switching mode power amplifier. Tripathi et al '089 does not show or suggest an extended interface as claimed by claim 8 and thus by virtue of dependency, claims 9, 11, and 16-17. Therefore and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed extended interface of claims 8-9, 11, and 16-17 and does not support a rejection of these claims under §102(b).

Claim 20 and 21in addition to being dependent on claim 1, recite further limitations, specifically and respectively, generation of a predistortion signal and various forms of linearizers, regarding the linearizer recited in claim 1 and these limitations are

not shown or suggested by Tripathi et al '089. Claims 82 and 83 are respectively analogous to claims 20 and 21 and further limit claim 77. Therefore and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed invention of claim 20-21 or claim 82-83 and thus does not support a rejection of these claims under §102(b).

Claim 28 further defines the switching mode power amplifier of claim 1 to have an adjustable output power. Claim 79 similarly limits claim 77. This is not shown or suggested by Tripathi et al '089. Therefore and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed invention of claim 28 or 79 and thus does not support a rejection of these claims under §102(b).

Claim 33-34 further defines the delta sigma modulator of claim 1 to be a multi-band band pass filters suitable for simultaneous operation on a plurality of frequency bands. This is not shown or suggested by Tripathi et al '089. Therefore and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed invention of claims 33-34 and thus does not support a rejection of these claims under §102(b).

Claim 36 – 37 as well as 38, recite specifics of supplying power to the power amplifier of claim 1. These limitations are not shown or suggested by Tripathi et al '089. Therefore and for these additional reasons, Applicant respectfully submits that Tripathi et al '089 does not show or suggest the claimed invention of claims 36-37 and thus does not support a rejection of these claims under §102(b).

Therefore and at least for the additional reasons noted above, Applicant respectfully submits that Tripathi et al '089 does show or suggest all limitations of claims 3-4, 8-9, 11, 20-21, 28, 33-37, 78-79, 82 and 83. Thus Applicant for these additional reasons respectfully requests the Examiner to reconsider and withdraw this rejection of claims 3-4, 8-9, 11, 20-21, 28, 33-37, 78-79, 82 and 83 under 35 U.S.C. 102(b) based on Tripathi, et al. (U.S. Patent No. 5,974,089).

b) Claims 16, 17 and 22-24 stand rejected under 35 U.S.C. 103(a) as being unpatentable over Tripathi, et al. (Fig. 2). These claims are each dependent on claim 1 and at least by reason of dependency on a claim that is believed to be allowable should likewise be allowable. Therefore, Applicant respectfully requests that the Examiner reconsider and withdraw this rejection of claims 16, 17 and 22-24 under 35 U.S.C. 103(a) based on Tripathi, et al.

c) Claims 5, 10, 12-15, 18-20, 26, 27, 29-32, 38, 80 and 81 are objected to as being dependent upon a rejected base claim. These claims have been objected to as dependent upon a rejected base claim but would be allowable if rewritten to include all limitations of the base and any intervening claims. Applicant appreciates and concurs with the Examiner's view that these claims define allowable subject matter. Applicant may wish to amend such claims if further discussions of the independent/intervening claims are not productive. Applicant further respectfully suggests that the Examiner may want to consider the claim numbers and

corresponding reasons for allowance as it appears that some claim numbers that the Examiner is using may be now be different.

Accordingly, Applicant respectfully submits that the claims, as amended, clearly and patentably distinguish over the cited references of record and as such are to be deemed allowable. Such allowance is hereby earnestly and respectfully solicited at an early date. If the Examiner has any suggestions or comments or questions, calls are welcomed at the phone number below.

Although it is not anticipated that any fees are due or payable, other than the two month extension of time separately authorized, the Commissioner is hereby authorized to charge any fees that may be required to Deposit Account No. **50-1147**.

Respectfully submitted,

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A DSP CONTROLLED ADAPTIVE FEEDFORWARD AMPLIFIER LINEARIZER

by

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B.A.Sc., University of British Columbia, 1993

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THE REQUIREMENTS FOR THE DEGREE OF

MASTER OF APPLIED SCIENCE

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of

Engineering Science

SIMON FRASER UNIVERSITY

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APPROVAL

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DEGREE: Master of Applied Science

TITLE OF THESIS: A DSP Controlled Adaptive Feedforward Amplifier Linearizer

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Date Approved: ____________________

ABSTRACT

Currently, the only available wideband (multiple MHz), accurate, amplifier linearization method is feedforward. Feedforward, though, requires automatic adaptation of key parameters for reliable distortion cancellation as operational and environmental conditions vary. In this thesis, previous analysis on adaptive feedforward linearization is extended to include an alternative placement of the adaptive signal cancellation coefficient. In contrast to previous analysis, this placement results in a non-quadratic error surface. Consequently, two available criteria for the optimization of the signal cancellation coefficient result in different optimal values. This result can have practical implementation consequences under certain operating conditions. A new analysis is presented that shows that various inaccuracies in the implementation of baseband correlation, such as frequency and phase offsets, filter mismatches, and incomplete image suppression, do not affect the final converged coefficient values. With a novel and appropriate use of DSP, a feedforward linearizer has been implemented with adaptation driven by easily computed gradient signals. This overcomes the difficulties, such as DC offsets at the output of analog mixers and masking of weak signals by stronger ones, that slow and/or cause incorrect convergence of many previously reported implementations. The result is 40 dB reduction of intermodulation spectra over a bandwidth of 7 MHz. Coefficient convergence occurs within 50 msec of start-up, and following a 6 dB change in input power, reconvergence occurs in 3 msec with no loss in distortion suppression.

ACKNOWLEDGEMENTS

Nothing is ever done alone; thus, I would like to thank certain people who helped me along the way with this project. Thanks to my Senior Supervisors, Jim Cavers and Paul Goud, who, I feel, were responsible for advancing my knowledge and confidence in the area of wireless communications to a new and hopefully much higher level than before. Deserving of thanks as well, is my wife, Irma, for her constant support the whole way through the project, from start to finish.

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1. INTRODUCTION

All wireless radio transmitters contain RF amplifiers which are nonlinear to some degree. The primary consequence of amplifier nonlinearities is the generation of intermodulation distortion (IMD) if the signal to be amplified has a non-constant envelope, such as for linear or multicarrier modulation formats. Not only does IMD corrupt the amplified signal itself, but more seriously it causes adjacent channel interference due to spectral regrowth. In interference limited systems such as cellular radio, strict limits are usually placed upon allowable intermodulation power in adjacent channels; consequently, some form of amplifier linearization is usually required. Several linearization techniques employed to combat IMD are feedback, predistortion, and feedforward. Adaptive feedforward is the scheme studied in this thesis.

1.1 Characterization of Amplifier Nonlinearities

Nonlinear RF amplifiers are characterized by measurement of their AM/AM (amplitude dependent gain) and AM/PM (amplitude dependent phase shift) characteristics. These measurements may be performed using a network analyzer in power sweep mode. Specifically, the gain and phase of the amplifier are measured at a single frequency as the input power level is varied. Not only are RF amplifiers nonlinear, but they also possess memory: the output signal depends on the current value of the input signal as well as previous values spanning the memory of the amplifier. If the reciprocal of the bandwidth of the input signal is much larger than the memory of the amplifier, as is the case for most RF amplifiers driven with narrowband signals, the amplifier can be modeled as memoryless

for that particular input signal. Thus, for complex baseband analytical and simulation purposes, the AM/AM and AM/PM measurements can be summarized in a single frequency-independent memoryless function, namely complex voltage gain

$$G(x) = g(x)e^{j\phi(x)} \tag{1}$$

The magnitude and phase of $G(x)$ are simply the measured gain and phase of the amplifier as functions of x—the instantaneous power in the input signal. For wideband signals, the memory of the amplifier becomes a significant fraction of the reciprocal of the bandwidth of the input signal; thus, it must be considered for accurate analytical and simulation studies. In this case, a single frequency-independent function is not sufficient to model the amplifier nonlinearity, and more powerful modeling techniques must be used such as a Volterra series approach. For the analysis presented in this thesis, though, the power amplifier is assumed to be memoryless.

Figure 1 shows the characteristics of a typical class AB RF power amplifier measured at a single frequency within the passband of the amplifier [1]. Note that the gain and input power are normalized such that saturation occurs at unity output power for unity input power. Consequently, normalized input power also represents input backoff from saturation. For example, 6 dB backoff corresponds to an input power of 0.25.

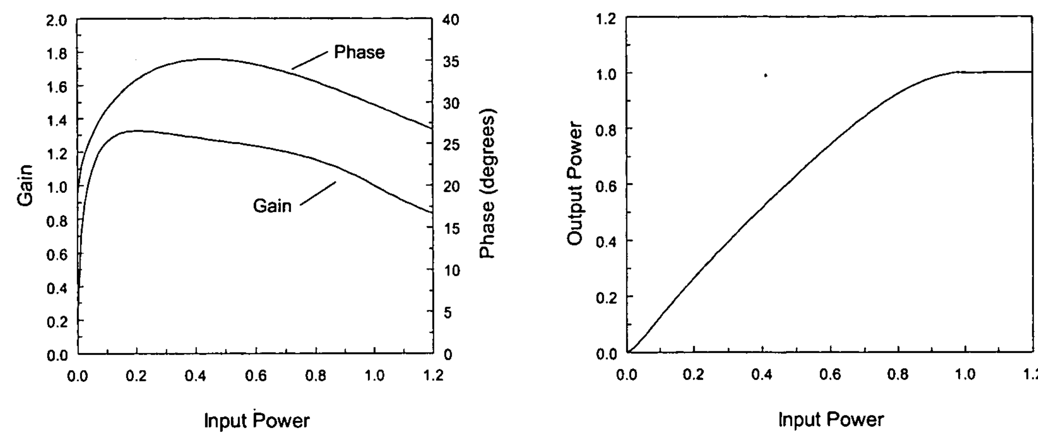

Figure 1. Measured characteristics of a typical class AB power amplifier

Evident in Figure 1 is a strong amplitude dependence of the amplifier gain and phase. Because the transistors in the amplifier are biased in class AB rather than class A, they cut off at low signal voltages. This causes the gain to fall off at low input power. The gain also rolls off at high input power due to saturation of the transistors. The variation in phase with input power is due to voltage-dependent device capacitances.

Figure 2 shows simulated power spectra of a narrowband $\pi/4$-DQPSK signal before and after amplification with the amplifier whose nonlinear characteristics are shown in Figure 1. For simulation purposes, $G(x)$ is represented by polynomials fitted to the measured gain and phase curves over the range of input powers extending from 0 to 1 (saturation); $G(x)$ is then extrapolated further into saturation. If $v_m(t)$ is the complex envelope of the amplifier input signal, then the complex envelope of the distorted amplifier output signal is given by $v_a(t) = v_m(t)G[|v_m(t)|^2]$. 35% rolloff root raised cosine filtering is used for the simulation which results in a peak-to-average power ratio of approximately 2.5 dB; input backoff is 3 dB. Note that frequency has been normalized by the symbol rate. As can be seen, the amplifier nonlinearity causes significant spectral regrowth.

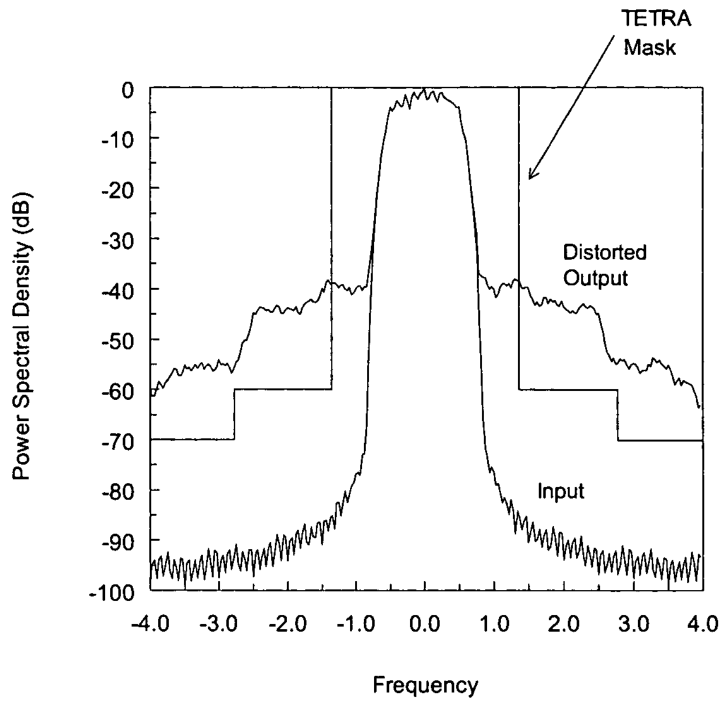

Figure 2. Simulated input and output power spectra for class AB amplifier with a $\pi/4$-DQPSK input signal

Regulatory bodies usually specify power spectral density (PSD) masks which define maximum allowable adjacent channel interference (ACI) levels. TETRA [2], for example, uses a $\pi/4$-DQPSK modulation format with 35% root raised cosine filtering at a symbol rate of 18 kHz; the channel spacing is 25 kHz. Adjacent channel protection is specified as 60 dB at 25 kHz, and 70 dB at 50 and 75 kHz. The corresponding spectrum mask is shown in Figure 2. Clearly, TETRA limits are exceeded; consequently, some form of linearization for this amplifier is required in order to conform to this standard.

1.2 Overview of Linearization Strategies

Perhaps the simplest method of achieving linear RF amplification without the use of additional hardware is output backoff of an existing class A amplifier such that it

operates completely in its linear region. Typically though, the backoff required to achieve linear operation is high (25 to 30 dB), and the resulting power efficiency is very low (1 to 2%). Also, for a fixed output power requirement, the cost of building an amplifier with output power rating 25 to 30 dB higher than necessary can be high. Moreover, the heat dissipation from the higher power amplifier may become a problem.

Several other techniques to achieve linear RF amplification have been developed. The most popular are feedback, predistortion, and feedforward, all of which make use of additional hardware. Making a choice of which linearization strategy to employ for a particular application involves tradeoffs of complexity, degree of IMD suppression, and bandwidth. The most prominent feedback scheme, namely Cartesian coordinate modulation feedback [3], has relatively low complexity, offers reasonable IMD suppression, but stability considerations typically limit the bandwidth to a few hundred kHz. Digital implementations of predistortion [4,5] have higher complexity than feedback, offer better IMD suppression, but again, possible bandwidths are low (up to a few tens of kHz) due to limited DSP computation rates. Reported implementations of analog predistortion [6], although potentially wideband, have modest complexity, but suffer from limited IMD suppression. In contrast to the above linearization techniques, feedforward [7,8,9] is currently the only linearization strategy that simultaneously offers wide bandwidth and good IMD suppression; the cost is relatively high complexity. Automatic adaptation of key parameters, though, as discussed in the next section, is essential for reliable distortion cancellation as operating conditions vary.

Wide bandwidth capability makes feedforward an attractive scheme for several applications. At cellular base stations, rather than using one amplifier per channel

followed by lossy high power combiners, it is more efficient to combine channels at low power and use a single wideband linearized amplifier for the resultant signal. Another potential use of feedforward is for emerging PCS applications, including wideband CDMA, in which the bandwidth requirements place feedback and predistortion out of the league of viable linearization schemes.

1.3 Development of Feedforward Linearization

In 1927, H.S. Black of Bell Telephone Laboratories invented the concept of negative feedback as a method of linearizing amplifiers for the Bell Telephone system [10]. Lesser known is that four years earlier, in the search for a linearization method, he invented the concept of feedforward. His idea for feedforward was simple: reduce the amplifier output to the same level as the input and subtract one from the other to leave only the distortion generated by the amplifier. Amplify the distortion with a separate amplifier and then subtract it from the original amplifier output to leave only a linearly amplified version of the input signal. Black's idea for negative feedback was spawned from his simple feedforward concept: feed an attenuated version of the amplifier output signal back to the input in anti-phase and combine it with the input signal. Use the same amplifier (rather than a separate amplifier as in feedforward) to amplify the difference signal thus producing a linearly amplified version of the input signal. The advantage of the feedback solution, due to the fact that it operated closed-loop, was that it was automatic and required no manual adjustment as operating conditions changed. Its disadvantage, of course, was its potential for instability.

The basic operating principles of a feedforward amplifier as shown in Figure 3 are now described. The feedforward configuration consists of two circuits, the signal cancellation circuit and the error cancellation circuit. The purpose of the signal cancellation circuit is to suppress the reference signal from the main amplifier (or PA) output signal leaving only amplifier distortion, both linear and nonlinear, in the error signal. Linear distortion, in contrast to nonlinear distortion described already, is due simply to deviations of the amplifier's frequency response from flat gain and linear phase [11]. Note that the feedforward technique can also compensate for memory effects, since distortion due to memory in the main amplifier is also included in the error signal and thus ultimately canceled in the linearizer output. The fact that the PA output can be decomposed into two components—a linear term and a distortion term—is discussed in Section 2.

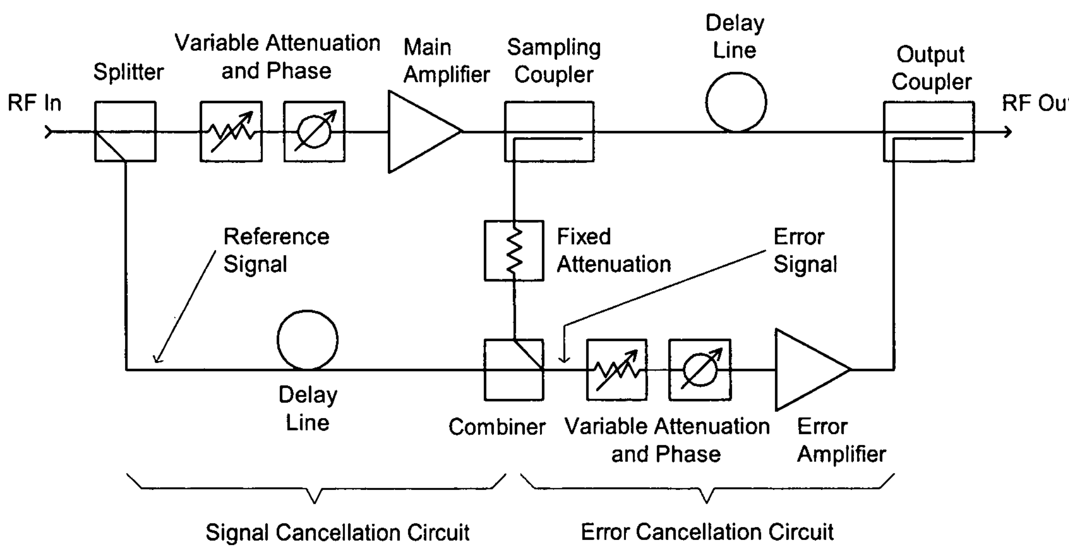

Figure 3. RF circuit configuration of a typical feedforward amplifier

In order to suppress the reference signal, the values of the sampling coupler and fixed attenuation are chosen to match the gain of the main amplifier so that the PA output signal is reduced to approximately the same level as the reference signal. The variable phase shifter ahead of the PA is then adjusted to place the PA output in anti-phase with the reference. The variable attenuation serves the fine tuning function of precisely matching the level of the PA output and the reference. The delay line in the reference branch, necessary for wide bandwidth operation, compensates for the group delay of the main amplifier by time aligning the PA output and reference signals before combining.

The purpose of the error cancellation circuit is to suppress the distortion component of the PA output signal leaving only the linearly amplified component in the linearizer output signal. In order to suppress the error signal, the gain of the error amplifier is chosen to match the sum of the values of the sampling coupler, fixed attenuator, and output coupler so that the error signal is increased to approximately the same level as the distortion component of the PA output signal. The variable phase shifter ahead of the error amplifier is then adjusted to place the error in anti-phase with the PA output. The variable attenuation, again, serves the fine tuning function of precisely matching the level of the error signal and the distortion component of the PA output. The delay line serves the same purpose as in the signal cancellation circuit. The error amplifier must be chosen such that it linearly amplifies the error signal while still providing the required output power, otherwise uncorrectable IMD shows up in the linearizer output. This usually dictates the use of a linear class A amplifier with sufficient backoff. Note that any bandwidth limit, manifested as incomplete distortion suppression, is imposed either by

imperfect delay matching or by linear distortion in the error amplifier, the variable attenuators/phase shifters, or the couplers and combiners.

The crux of the proper operation of the feedforward circuit is the proper adjustment of the attenuation and phase in the signal and error cancellation circuits such that good IMD suppression is maintained over time. Variations of component characteristics with temperature and time as well as changes in operating conditions such as input power level and supply voltage all necessitate readjustment. For these reasons Black, himself, essentially abandoned feedforward in favour of feedback. He found, using vacuum tube amplifiers, that "every hour on the hour—24 hours a day—somebody had to adjust the filament current to its correct value. In doing this, they were permitting plus or minus 1/2- to 1-dB variation in amplifier gain, whereas, for my purpose, the gain had to be absolutely perfect. In addition, every six hours it became necessary to adjust the B battery voltage, because the amplifier gain would be out of hand. There were other complications too, but these were enough!" [10]. Even with modern solid state amplifiers, changes with temperature and time are still significant enough, with respect to the accuracy requirements of the feedforward circuit, to necessitate adaptation.

After its invention in 1923, feedforward was essentially ignored until Seidel, also at Bell Laboratories, investigated the use of feedforward for microwave frequency TWT amplifiers in the late sixties and early seventies [7]. Seidel constructed a feedforward amplifier which achieved distortion suppression of 38 dB over a 20 MHz band. The setup employed an automatic control scheme for the variable attenuation and phase in the error cancellation circuit. The control scheme was based on driving a mechanical attenuator/phase shifter with an error signal derived by comparing the amplitude and phase

at two different points in the error cancellation circuit of a pilot tone inserted after the main amplifier. No control scheme was utilized for the variable attenuation and phase in the signal cancellation circuit. In this way he was able to maintain time independent distortion suppression over a period of several months.

Several patents concerned with adaptive feedforward systems then started to appear in the mid-eighties, and many more appeared in the early nineties. These patents deal with two general methods of adaptation both with or without the use of pilot tones, namely adaptation based on power minimization [12,13] and adaptation based on gradient signals [14,15]. The control scheme for the former attempts to adjust the attenuation and phase in the signal cancellation circuit in such a way to minimize the measured power of the error signal in the frequency band occupied by the reference signal. Minimum power in the error signal is equivalent to suppression of the reference signal. The attenuation and phase in the error cancellation circuit are adjusted in such a way to minimize the measured power of the linearizer output signal in a frequency band occupied only by distortion. Minimum power in the output signal is equivalent to suppression of distortion. Once the optimal parameters are found, deliberate misadjustment is required over time to assess whether or not the respective powers are indeed still minimized. This deliberate perturbation periodically reduces IMD suppression—an undesirable side effect.

Adaptation using gradient signals is based on continually computing estimates of the gradient of a 3-dimensional power surface which depends on two parameters—the variable attenuation and phase in either the signal or error cancellation circuits. The gradient signal is then used to adapt the parameters in each circuit always in a direction towards the global minimum of the surface. The surface for the signal cancellation circuit

is the power in the error signal; the power is minimized when the reference signal is completely suppressed, leaving only the amplifier distortion in the error signal. The surface for the error cancellation circuit is the power in the linearizer output signal; the power is minimized when the distortion is completely suppressed from the PA output signal. The advantage of adaptation based on gradient signals over that based on power minimization, is that since the gradient signals are continually computed, the control scheme constantly searches for the optimum operating point. No algorithm for deliberate misadjustment is required.

Adaptation using either of the above methods plus pilot tones is based on inserting a pilot both at the input to the feedforward linearizer as artificial signal and at the output of the main amplifier as artificial distortion. The control scheme for the signal cancellation circuit either attempts to null the first pilot tone in the error signal if using the power minimization approach, or uses it to derive a gradient signal if using the gradient approach. The same is true for the second pilot tone, except the observation point is the linearizer output. When both pilots are canceled, so is the amplifier distortion. As is the case for other components in a communication system, it is desirable to avoid pilot tones, if possible, and use traffic signals only.

Other than in the patent literature, very little has been published on implementations or analysis of adaptive feedforward linearizers since Seidel's work. Two publications of note on adaptive feedforward linearization, though, are by Cavers [16] and Narahashi and Nojima [17]. Cavers' work is the first published analysis of adaptation behaviour of a feedforward linearizer and is intended as a benchmark and analytical framework for others developing such linearizers.

Narahashi and Nojima report an implementation of an adaptive feedforward linearizer for multicarrier signals with adaptation based on a power minimization technique using pilot tones. A pilot tone is inserted at the output of the main amplifier and its level is measured in the linearizer output signal by means of a narrowband energy detector. A microprocessor is used to adjust the attenuation and phase in the error cancellation circuit using a perturbation technique. Adaptation in the signal cancellation circuit is performed using the same method, except that one of the carriers is used as the pilot signal, rather than an explicit pilot tone. With this setup, it is reported that 30 dB distortion improvement is obtained in a stable manner for a 100 W, 1.5 GHz, GaAs-FET power amplifier.

1.4 Project Goals

Based on an increasing need for a wide bandwidth, adaptive linearization technique and a lack of published work on adaptive feedforward, it was decided to implement, in contrast to [17], a gradient driven adaptive feedforward linearizer without the use of pilot tones. Gradient adaptation was selected because of the advantages it offers over the power minimization technique as discussed above. Based on available equipment, a 5 Watt, 815 MHz, class AB amplifier was chosen as the main amplifier.

A number of patents, e.g. [14], propose a gradient adaptation technique that relies on analog bandpass correlation requiring the mixing of two modulated RF signals. This method, elaborated on later, suffers from accuracy problems such as mixer DC offsets and the generation of additional IMD—both highly undesirable effects. To overcome these problems, the current work demonstrates a novel and appropriate use of DSP to perform

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A DSP Controlled Adaptive Feedforward Amplifier Linearizer

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Abstract—**Currently, the only available wideband (multiple MHz) amplifier linearization method is feedforward. Feedforward, though, requires automatic adaptation of key parameters for reliable distortion cancellation as operating conditions vary. With a novel and appropriate use of DSP, we have implemented a feedforward linearizer with adaptation driven by easily computed gradient signals. This overcomes the difficulties, such as DC offsets at the output of analog mixers and masking of weak signals by stronger ones, that slow and/or cause incorrect convergence of many previously reported implementations. The result is an average of 40 dB reduction in intermodulation spectra over a wide bandwidth, with extremely fast tracking.**

INTRODUCTION

RF power amplifiers (PAs) generate intermodulation (IM) distortion if they carry linear modulation or several frequency channels at once. Since IM distortion appears as interference in other channels, amplifier linearization is an essential element of a PCS. Of the known linearization strategies [1-5], only feedforward can provide good IM reduction over an operating bandwidth of tens of MHz, and it does so through its use of inherently wideband analog technology. The feedforward method, though, is very sensitive to parameter changes due to varying operating conditions such as input power level, supply voltage and temperature. Adaptation is thus essential.

Two significant problems have compromised many previous adaptation structures, e.g. [6]. One of them is the generation of a gradient signal by bandpass correlation of RF signals to produce a DC result. With analog mixing of the RF signals, any DC offset in the result causes convergence to incorrect parameter values. A second problem is masking of weak signals by strong ones in one gradient calculation, causing extremely slow convergence. This paper demonstrates a novel and appropriate use of DSP to solve both problems regardless of the bandwidth of the carried signals.

THE FEEDFORWARD STRUCTURE

Basic Feedforward Operation

Fig. 1 shows that a feedforward linearizer consists of cascaded signal cancellation and error cancellation circuits. In the signal cancellation circuit, the RF input signal $v_m(t)$ is split into amplification and reference branches. In the amplification branch, the signal is attenuated and shifted in phase by controllable values, and then fed to the PA. The PA output signal $v_a(t)$ is sampled with a coupler and attenuated to reduce its level to that of the reference signal where the two are combined in antiphase. With appropriate selection of the controllable attenuation and phase shift, the reference signal is canceled, leaving an error signal $v_e(t)$ which is an attenuated version of the IMD generated by the PA. In the error cancellation circuit, the error signal is attenuated and shifted in phase, again by a controllable value, and amplified in an error amplifier. When it is injected back into the main signal path, the IMD is canceled, leaving only a linearly amplified version $v_o(t)$ of the RF input. Delay lines in each circuit are necessary to compensate for the group delay of the PA and the error amplifier.

The crux of the operation of the adaptive feedforward linearizer is proper adjustment of the attenuation and phase parameters in each circuit and selection of a highly linear error amplifier so that no additional (uncorrectable) IMD is introduced in the linearizer output. These issues will be discussed in more detail in the subsequent sections.

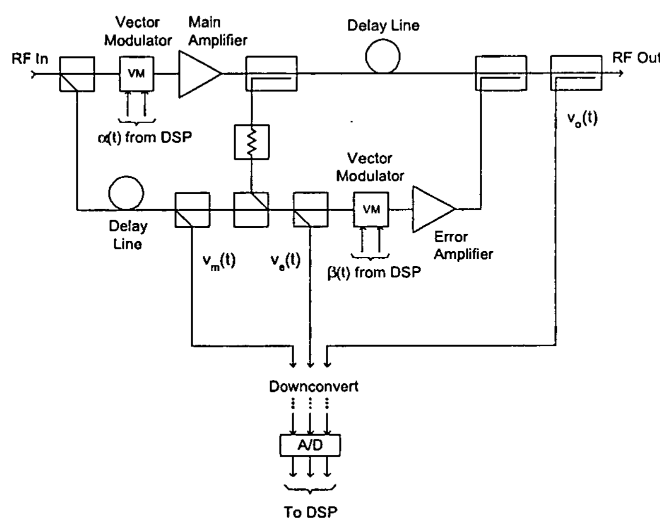

Fig. 1. Adaptive feedforward amplifier linearizer

Principles of Adaptation

A simplified baseband equivalent (Fig. 2) of the feedforward linearizer is useful for analyzing adaptation. Signals are replaced by their complex envelopes. Adaptive complex coefficients α and β represent the attenuation and

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phase shift introduced in both the signal and error cancellation circuits respectively. The power amplifier is modeled as a memoryless nonlinearity whose AM/AM and AM/PM conversion is summarized by its complex voltage gain $G(x)$, where x denotes instantaneous power. The PA output $v_a(t)$ is therefore

$$v_a(t) = \alpha v_m(t) G\left(\left|\alpha v_m(t)\right|^2\right) \qquad (1)$$

which consists of a linearly amplified term $c_0 \alpha v_m(t)$ and a distortion term $v_d(t)$. The complex coefficient c_0 is the first coefficient of the power series expansion of the nonlinear function $G(x)$. For this analysis, we also assume that delay matching is accurate and that the error amplifier is perfectly linear.

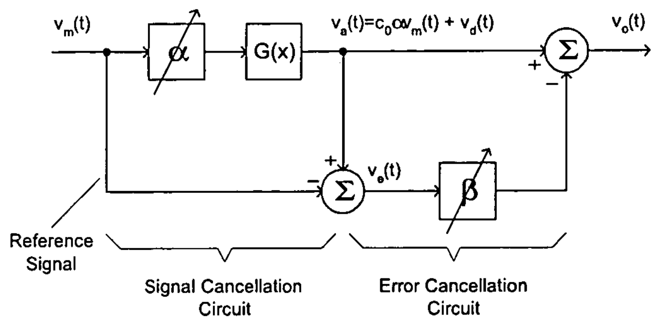

Fig. 2. Complex baseband equivalent model

A detailed analysis of gradient based adaptation without pilot signals, including accuracy requirements and convergence time and jitter, is contained in [7], and is summarized here. Adaptation of α, based on the linear estimation of $v_a(t)$ with basis $v_m(t)$, is designed to minimize P_e, the average power of the error signal. When P_e is minimized, the error signal is simply the IMD introduced by the PA, $v_d(t)$, with the reference signal completely canceled. If the coefficient α were placed in the lower (reference) branch, making $v_e(t) = v_a(t) - \alpha v_m(t)$, then P_e would be a quadratic function of α with a well defined minimum. A suitable gradient signal would be the covariance of the estimation error and the reference signal, for which a noisy, but unbiased estimate is

$$D_\alpha(t) = v_e(t) v_m^*(t) \qquad (2)$$

When α is adjusted properly, $E[D_\alpha(t)] = 0$, since the optimal linear estimate occurs when the estimation error and the basis of the estimate are uncorrelated. This suggests the following algorithm for the adaptation of α implemented as a first order circuit

$$\alpha(t) = K_\alpha \int_0^t D_\alpha(\tau) d\tau \qquad (3)$$

where K_α controls the time constant of the adaptation. For our placement of α in the upper (amplification) branch instead, it is shown in [8] that an identical gradient signal and adaptation algorithm cause convergence to the optimum value, despite the interaction of the coefficient and the PA.

The adaptation of β proceeds in a similar manner except that the basis of the estimate of $v_a(t)$ is $v_e(t)$ and the estimation error is $v_o(t)$. The gradient signal for the adjustment of β is

$$D_\beta(t) = v_o(t) v_e^*(t) \qquad (4)$$

and the algorithm for adaptation is

$$\beta(t) = K_\beta \int_0^t D_\beta(\tau) d\tau \qquad (5)$$

where K_β is analogous to K_α.

Problems in Adaptation

There are two main problems in conventional feedforward adaptation. The first appears if the gradients (2) and (4) are implemented by bandpass correlation of the corresponding RF signals to produce a lowpass result. DC offsets and $1/f$ noise in the mixers cause convergence to incorrect values, weakening the degree of IM suppression. The second is that the gradient (4) relies on mixing of $v_e(t)$ with the weak IM component in $v_a(t)$. The much stronger signal component in $v_a(t)$ acts as noise, forcing a very long time constant and thus slow convergence. We have solved both problems—the first, by use of DSP to perform correlation, and the second, by use of a filter to suppress the desired signal component in $v_o(t)$.

IMPLEMENTATION

Equation (3) is implemented in DSP by use of the familiar LMS algorithm

$$\alpha(n) = \alpha(n-1) + \delta D_\alpha(n) \qquad (6)$$

where the parameter δ is the step size. Equation (5) is implemented in an analogous manner. Fig. 1 shows the points in the feedforward circuit where the appropriate RF signals are split to enable downconversion and sampling before recovering the required complex envelopes in DSP.

The real and imaginary components of $\alpha(n)$ and $\beta(n)$ are used directly as the control signals for the vector modulators (VMs) in the signal and error cancellation circuits which realize the required attenuation and phase shift. Both VMs have highly linear RF input/output characteristics—an important property, especially in the error cancellation circuit. The attenuation and phase shift provided by the VMs varies monotonically with the control voltages; this ensures convergence of α and β to the correct values.

For wide bandwidth operation, the gradient signals for the adaptation of α and β cannot be computed in one step due to limited sampling rates available for DSP. Thus, the gradient computations must be performed in subbands spanning selected spectral regions of the signals and the result from each subband summed. This method is validated by replacing the time domain integrals in equations (3) and (5) with the corresponding frequency domain integrals using Parseval's theorem. Various subbands may be selected by employing a

two- or three-step downconversion process in which the frequency of the first oscillator controls which portion of the signal is placed in the narrow window of the first IF filter; the second and/or third oscillators remain fixed.

The main amplifier in Fig. 1 is a 5 Watt, 815 MHz, class AB PA driven by a narrowband $\pi/4$-DQPSK modulated signal centred about 815 MHz. The data signal employs root raised cosine pulse shaping with 35% rolloff at a symbol rate of 20 ksym/s. A narrowband, rather than wideband, data signal is used so that the gradient computation may be performed in only one step, rather than several steps as described above. The error amplifier is a 5 W class A amplifier which operates with sufficient backoff to ensure linear operation.

Since $v_m(t)$, $v_e(t)$, and $v_o(t)$ are bandpass signals, the DC offsets generated by the mixers in the downconversion process can be avoided by centering the signals about an IF low enough to be sampled, but high enough to ensure no spectral occupancy at DC, and then accurately recovering the corresponding complex envelopes in DSP. In the present implementation, this is achieved using a sampling rate of f_s = 150 kHz and an IF of $f_s/4$. Note that the same oscillators are used to downconvert each of the three signals $v_m(t)$, $v_e(t)$, and $v_o(t)$.

For the adaptation of α, the complex envelopes of $v_m(t)$ and $v_e(t)$ are recovered using complex bandpass FIR filters of length 16 centred about $f_s/4$ which provide 40 dB image suppression. The filter stopband extends from $f_s/2$ to f_s. The high stopband attenuation is achieved with a short filter by allowing the transition bandwidth to be wide. This is acceptable, since the spectral region of interest for the correlation of $v_m(t)$ and $v_e(t)$ is that occupied by the reference signal only. No sample-by-sample derotation of the filter outputs is necessary since the constellation rotations are canceled by the complex conjugate multiplication performed in the gradient calculation. For the same reason, any frequency offset in the downconverted signals from the IF of $f_s/4$ is canceled.

The adaptation of β is implemented in a similar fashion with one notable difference. The masking problem is solved by modifying the filter used to recover the complex envelope of $v_o(t)$ by introducing a notch in the bandpass frequency response to suppress the signal component. The resulting filter is a complex bandstop FIR filter of length 53 which provides 60 dB attenuation of the signal component and 40 dB image suppression. In contrast to the filters used for the adaptation of α, the transition bandwidths for the bandstop filter must be narrow, since the spectral region of interest for the correlation of $v_o(t)$ and $v_e(t)$ is that occupied by the distortion outside the band of the desired signal. A similar complex bandpass filter to those used for the adaptation of α is used for $v_e(t)$, except its length is 53 to match the length of the bandstop filter used for $v_o(t)$, and to ensure narrow transition bandwidths.

To reduce development time, the DSP used for the adaptation of α and β is TI's TMS320C30 floating point processor. Due to processing power constraints, the filter outputs for the adaptation of α and β are decimated by a factor equal to the corresponding filter lengths—16 for α and 53 for β. The effect of this simplification is a longer adaptation time constant as shown in the following section. Note that the choice of the filter lengths for the decimation factors simplifies the DSP code; smaller factors could be used. Moreover, with a slightly faster DSP, decimation would not be necessary at all.

RESULTS

The convergence behavior of α and β is illustrated in plots of the control voltages for the vector modulators in the signal and error cancellation circuits (Figs. 3 & 4). Starting from zero, the initial convergence time of α is less than 0.3 sec, and the reconvergence time due to a 6 dB drop in input power at time t ≈ 5.55 is less than 50 msec. As can be seen in Fig. 4, β is not affected by the step change in input power, thus no loss in IM suppression occurs during this time. The initial convergence time of β, approximately 2.5 sec, is longer than that for α, but once converged, β does not have to adapt to changes in operating conditions as quickly as α. Note that because of the decimation involved in the DSP implementation of the LMS algorithm discussed previously, the convergence time for α is slowed by a factor of 16, and for β by a factor of 53. Thus, if no decimation is required, the initial convergence times of α and β would drop to approximately 20 and 50 msec respectively, and the reconvergence time of α would drop to approximately 3 msec.

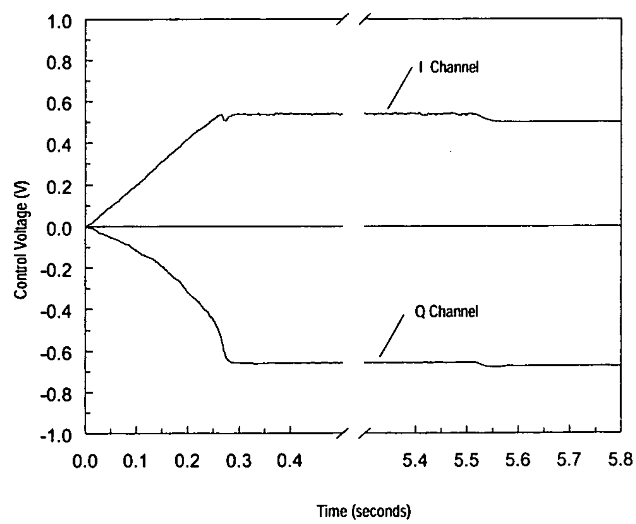

Fig. 3. Convergence of α

Fig. 5 illustrates approximately a 30 dB reduction in intermodulation spectra achieved by the adaptive feedforward linearizer for an input narrowband data signal. The resulting linearizer output spectrum is virtually identical to the spectrum of the input signal. The corresponding PA output power is +34 dBm, which is approximately 3 dB below the amplifier's output 1 dB compression point. Note that the plot of the output spectrum without linearization is obtained by allowing α to

converge to its optimal value which determines the PA operating point, and then disconnecting the error cancellation circuit from the output coupler so that the amplifier distortion is not suppressed.

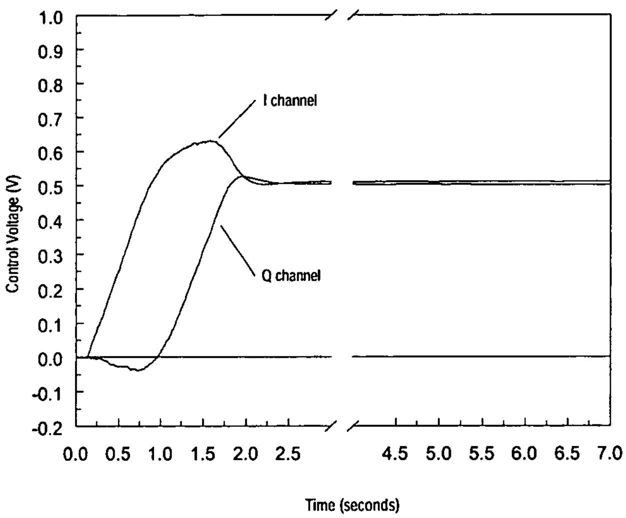

Fig. 4. Convergence of β

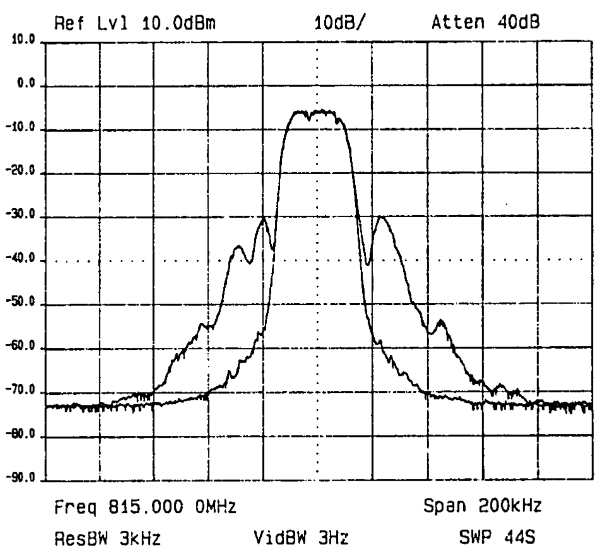

Fig. 5. Output spectra for power amplifier with and without linearization for a narrowband data signal

Fig. 6 demonstrates that the feedforward linearizer is capable of wide bandwidth operation. A single tone at 812.5 MHz (2.5 MHz offset from band center) is combined with the narrowband data signal at the linearizer input. Note that the adaptation algorithm is still based on the narrowband data signal centred at 815 MHz as before; in effect, the added tone is transparent. At the output of the power amplifier, intermodulation products appearing at multiples of 2.5 MHz offset from band center are indeed canceled by up to 40 dB evident from the residual IM product at 817.5 MHz. Further tests show 40 dB cancellation for tone offsets up to 3.5 MHz, thus the effective bandwidth is double this, i.e. 7 MHz. The effective bandwidth for 30 dB cancellation is measured as 10 MHz.

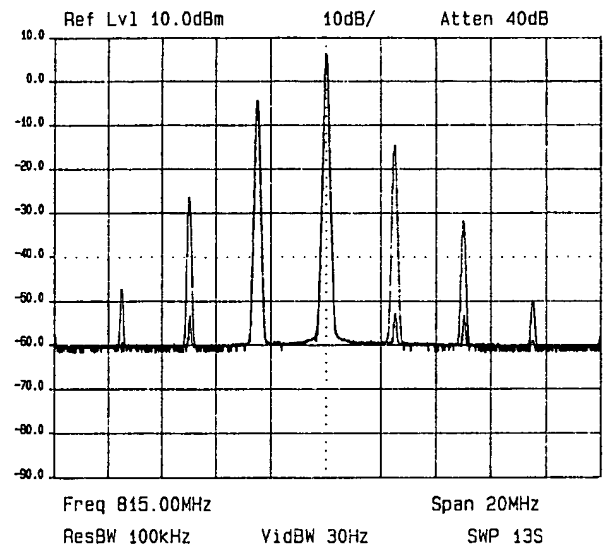

Fig. 6. Output spectra for narrowband data signal plus tone

CONCLUSIONS

We have implemented a gradient driven adaptive feedforward amplifier linearizer with a novel use of DSP that overcomes the problems of mixer DC offsets and masking of strong signals by weaker ones that compromise other previously proposed analog implementations. The result is a 40 dB reduction in IM spectra over a bandwidth of at least 7 MHz with extremely fast tracking. Initial convergence of the adaptation algorithm occurs in approximately 50 msec, and following a sudden reduction of signal level by 6 dB, it reconverges in approximately 3 msec, during which time there is no loss of IM suppression.

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